DEMODULATION METHOD AND DEMODULATOR

FOR CPFSK-MODULATED SIGNALS

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Background of the Invention:

Field of the Invention:

The invention lies in the telecommunications field and relates, more specifically, to a method and a device for estimating a sequence of input data symbols of a CPFSK-modulated data signal transmitted via a faulty channel.

A multiplicity of digital types of modulation are known which are based on amplitude shift keying (ASK), frequency shift keying (FSK), or phase shift keying (PSK), and mixtures of the same. What are termed CPM types of modulation (CPM: Continuous Phase Modulation) with continuous phase are used in digital communication systems for reasons of frequency economy. FSK with continuous phase is noted as CPFSK (Continuous Phase FSK). An example of this is Gaussian minimum shift keying (GMSK), which is used, for example, in the pan European mobile radio standard GSM (Global System for Mobile Communications).

The originally non-linear CPFSK modulation can be described approximately as linear modulation. The linear approximation on which this property is based is specified in the article

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entitled "Exact and Approximate Construction of Digital Phase
Modulations by Superposition of Amplitude Modulated Pulses
(AMP)" by Pierre A. Laurent, IEEE Trans. Commun., Vol. COM-34
(1986), pages 150-160.

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The demodulation of a CPFSK -modulated data signal transmitted
via a distortionless and interference-free channel is
described, for example, in the book entitled
"Nachrichtenübertragung" ["Telecommunication"] by K. D.
Kammeyer, B. G. Teubner Verlag, Stuttgart 1996, chapter
12.1.5., pages 422 and 423. Given a modulation index η of 0.5
or a multiple of 0.5, all that is required is to sample the in
phase and quadrature branches of the received signal (because
of the 90° phase offset between these branches), and compare
the samples obtained with the corresponding complex-value
representations of the CPFSK replacement symbols on which the
linear approximation is based. Determined, as input data
symbol actually sent, from the possible input data symbols is
the one whose complex-value replacement symbol comes closest
(assuming a known initial phase) to the two measured samples
(real and imaginary parts).

This coherent demodulation method for CPFSK signals can be
generalized without problems to rational modulation indices
η = m1/m2 (m1 and m2 being integral), since a finite number of

states of the replacement symbols always exists in the case of
rational modulation indices, and so the demodulation and
(continue to) be performed by a comparison of the samples with
the aid of the finite modulation alphabet of the replacement

5   symbols.

However, there is no longer any finite modulation alphabet of
the replacement symbols for non-rational modulation indices η.
As a result of this, the conventional method for coherent

10   CPFSK demodulation can no longer be carried out.

A similar statement holds for a distorting transmission
channel wherein intersymbol interference (ISI) during
demodulation must be taken into account: as long as a rational

15   modulation index η is used, the demodulation of the data
signal transmitted via the distorted channel, for example by
means of the known Viterbi algorithm (VA), succeeds. In the
case of VA (which will be explained in more detail later) a
state diagram (what is termed a trellis diagram) is created in

20   accordance with the power of the replacement symbol alphabet
and the extent to which the ISI is taken into account (that is
to say the length L of the channel memory), and a transmitted
data sequence is determined as the shortest path through the
trellis diagram. If the modulation index η is an irrational

25   number, the realization or implementation of a VA is

impossible, however, because an infinitely large number of trellis states would have to be taken into account.

A general problem (that is to say one not limited to the CPFSK
5   form modulation considered here) in the demodulation of a data signal transmitted via a faulty, ISI-affected channel resides in the presence of unknown parameters in the receiver. In principle, the unknown parameters can be both channel parameters (for example the channel pulse responses) and
10   function parameters of a transmitter. Various methods are known for estimating unknown parameters (with reference to the transmitter and/or channel states) in the VA demodulation in the receiver. The classical mode of procedure is represented by DFE (Decision Feedback Equalization) techniques. In this
15   case, the same, estimated (incompletely known) parameters are used for all possible transition switching in two consecutive time steps in the trellis for which transition metrics are to be calculated, and these universal parameters are estimated with the aid of a fed back "provisional" sequence, present
20   with a delay of a few time steps, of demodulated data symbols which - depending on the delay selected - already corresponds more (long delay) or less (short delay) effectively to the sequence of input data symbols for which a search is being made. No further detail will be given here on the known DFE
25   techniques, although it may be mentioned that a principle problem of this technique consists in estimating unknown

parameters by using fed back "provisional" result data in conjunction with selection of a short delay is critical, because the error rate on the provisional result data can then still be too high and can become unsatisfactory on selection

5 of a longer delay time, because then the estimated parameters (for example channel pulse responses) are already obsolete at the instant of the estimation.

Another possibility for carrying out the VA given the presence

10 of unknown parameters is described in U.S. Patent No. 5,432,821. That method, denoted as Per-Survivor Processing (PSP) is based on the idea of undertaking for each path followed in the trellis diagram and progressed step by step a separate or individual estimation of the unknown parameters

15 based on variables which are assigned only to this considered path - and are therefore already calculated at the instant of the estimation - and to take this estimate into account when calculating the transition metrics. This mode of procedure, which is explained in U.S. Patent No. 5,432,821 for the

20 example of estimating channel parameters, has the advantage that there is no need for any time-delayed sequence of provisionally decided result data to be calculated and fed back.

Summary of the Invention:

The object of the present invention is to provide a demodulation method and demodulator for CPFSK-modulated signals which overcome the above-noted deficiencies and disadvantages of the prior art devices and methods of this general kind, and which enables demodulating a sequence of input data symbols of a CPFSK-modulated data signal, transmitted via a faulty channel, which permits a non-rational modulation index to be used at the transmitter end and, in particular, also permits deficient demodulation in the case of the presence of unknown parameters.

With the above and other objects in view there is provided, in accordance with the invention, a method of estimating a sequence of input data symbols of a CPFSK-modulated data signal, transmitted via a faulty channel, which comprises:

repeatedly executing an ACS operation and thereby calculating a transition metric value with reference to a transition from an output state at a time step n to a target state at a time step $n+1$, by

determining a first estimated value for a replacement symbol, occurring in a linear approximation of the CPFSK, with reference to an $n-1$th time step, and

taking the first estimated value into account in a
calculation of the transition metric value.

In accordance with an added feature of the invention, the
5   first estimated value is determined by using the first
estimated value determined in a preceding time step.

In accordance with an additional feature of the invention, the
first estimated value is separately determined for each output
10   state, specifically on a basis of decisions on input data
symbols taken on the path $(P(z_n^i))$ leading to the respective
output state (z_n^i).

In accordance with another feature of the invention, the first
15   estimated value is determined using the equation
$\hat{a}_{n-1}^{(i)} = \hat{a}_{n-2}^{(i)} \exp\left\{j\pi\eta d_{n-1}^{P(z_n^i)}\right\}$, whereby $\hat{a}_{n-1}^{(i)}$ and $\hat{a}_{n-2}^{(i)}$ are the first estimated
values for the n-1th and n-2th replacement symbol,
respectively, relating to the output state with index i, $d_{n-1}^{P(z_n^i)}$
is an input data symbol, decided in the receiver, with
20   reference to a path leading to the respective output state z_n^i,
and η denotes the modulation index.

In accordance with again an added feature of the invention, a
second estimated value is determined for a phase correction of

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a reconstructed signal value determined for calculating the transition metric value by using the first estimated value, and the second estimated value is considered in calculating the transition metric value.

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In accordance with again an additional feature of the invention, a second estimated value is separately determined for each output state, specifically on a basis of decisions on input data symbols taken on the path $(P(z_n^i))$ leading to the respective output state (z_n^i).

In accordance with a further feature of the invention, the step of determining the second estimated value comprises calculating a phase difference between a reconstructed signal value and a value of the received data symbol.

With the above and other objects in view there is also provided, in accordance with the invention, a device for estimating a sequence of input data symbols of a CPFSK-modulated data signal transmitted via a faulty channel, comprising:

a device configured to carry out ACS operations;

a calculating unit for calculating a transition metric value

with reference to a transition from an output state at a time

step n to a target state at a time step $n+1$, and

estimating means for determining a first estimated value for a

5 replacement symbol, occurring in a linear approximation of a

CPFSK, with reference to an n-1th time step, the estimating

means being connected to the calculating unit for

communicating the first estimated value to the calculating

unit.

In accordance with yet an added feature of the invention, the

estimating means is configured to carry out the determination

of the first estimated value by using the first estimated

value determined in a preceding time step.

In accordance with yet an additional feature of the invention,

the estimating means comprises a multiplicity of calculating

sections, and

each calculating section is configured to carry out a separate

20 calculation of a first estimated value for the n-1th

replacement symbol on the basis of decisions taken on input

data symbols for a path $(P(z_n^i))$ leading to the respective

output state (z_n^i) under consideration.

In accordance with yet another feature of the invention, each calculating section is configured to carry out the calculation of the first estimated value using the equation

5    $\hat{a}_{n-1}^{(i)} = \hat{a}_{n-2}^{(i)} \exp\left\{j\pi\eta d_{n-1}^{P(z_n^i)}\right\}$,   where $\hat{a}_{n-1}^{(i)}$ and $\hat{a}_{n-2}^{(i)}$ respectively are the first estimated values for the n-1th and n-2th replacement symbols relating to the output state of index i, $d_{n-1}^{P(z_n^i)}$ is an input data symbol decided in the receiver with reference to the path leading to the output state (z_n^i) under consideration,

10   and η denotes the modulation index.

In accordance with yet a further feature of the invention, the estimating means are first means and the device further comprises second means for determining a second estimated

15   value for a phase correction of a reconstructed signal value determined for calculating the transition metric value by using the first estimated value.

In accordance with a concomitant feature of the invention,

20   the second means comprise a multiplicity of calculating sections; and

each calculating section is configured to carry out a separate calculation of a second estimated value for the n-1th

replacement symbol on a basis of decisions taken on input data symbols for the path $(P(z_n^i))$ leading to the respective output state (z_n^i) under consideration.

5   An important aspect of the invention consists in using a first estimated value for the placement symbol occurring in the preceding time step n-1 during the linear approximation of the CPFSK in order to calculate the transition metric value for a transition between the time steps n and n+1. Thus, in the time

10   step n the state of the transmitter in the time step n-1 is estimated and the calculation of the reconstructed signal value is carried out on the basis of this estimated transmitter state. This mode of procedure according to the invention is also denoted below as TST (Transmitter State

15   Tracking). Since the transmitter state (lying one time step back) (that is to say the appropriate replacement symbol) is estimated, the presence of a finite modulation alphabet is not required. For a method according to the invention is therefore also suitable for using irrational modulation indices.

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The determination of the first estimated value is preferably performed by using the first estimated value determined in the preceding time step, that is to say it is possible to speak of an incremental determination of the first estimated value.

25

Furthermore, it is preferred to determine a second estimated value, which represents a phase correction of the reconstructed signal value required for calculating the transition metric value. This opens up the possibility of

5   carrying out the TST demodulation method even without exact knowledge of parameters which effect a phase error in the reconstructed signal value.

A first example for such a parameter is the modulation index

10   η. If the modulation index η in the receiver is not known, or not known exactly, this has the consequence that a phase shift occurs between the actual transmitter state and the directed transmitter state estimated in accordance with the TST method. Since the corrected transmitter state features in the

15   calculation of the reconstructed signal value and thus in the calculation of the transition metric value, this would impair the demodulation performance. An efficient signal demodulation is rendered possible by the determination of the second estimated value for a phase correction of the reconstructed

20   signal value, even in the case of a modulation index η, which is not known, or not known exactly, at the receiver end.

A further example for such a parameter is - for the case that the frequency modulation is undertaken relative to a carrier

25   frequency and, at the receiver end, the shift by the carrier

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frequency is nullified by downward mixing - is a frequency

offset between the carrier frequency, used for upward mixing,

of a transmitter and the carrier frequency, used for downward

mixing, of the receiver. This frequency offset, which can also

5   be effected by a drift of one or both mixing frequencies,

likewise acts as a phase error in the reconstructed signal.

Thus, in addition to elements of the modulation index in the

receiver it is also possible by means of the concept according

to the invention to take into account frequency offsets and/or

10 frequency drifts between the transmitter and the receiver.

This second estimated value is preferably determined for the

nth time step by calculating the phase difference between the

reconstructed signal value and a value of a received data

15 symbol. It is possible thereby to update the second estimated

value in each time step and to take account of it in

calculating the transition metric value for the corresponding

path.

20 Other features which are considered as characteristic for the

invention are set forth in the appended claims.

Although the invention is illustrated and described herein as

embodied in a demodulation method and a demodulator for CPFSK-

25 modulated signals, it is nevertheless not intended to be

limited to the details shown, since various modifications and

structural changes may be made therein without departing from the spirit of the invention and within the scope and range of equivalents of the claims.

5   The construction and method of operation of the invention, however, together with additional objects and advantages thereof will be best understood from the following description of specific embodiments when read in connection with the accompanying drawings.

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Brief Description of the Drawings:

Fig. 1 is a block diagram explaining the principle of the structure of a digital transmission system;

15   Fig. 2 is a block diagram explaining the mode of operation of a binary CPFSK modulator according to the prior art;

Fig. 3 is a block diagram explaining the basic structure of a coherent CPFSK modulator;

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Fig. 4a is a vector or signal space representation of the possible replacement symbols in the case of CPFSK with a modulation index $\eta = 0.5$;

Fig. 4b is a vector or signal space representation of the possible replacement symbols in the case of CPFSK with an irrational modulation index η;

5 Fig. 5 is an equivalent circuit diagram of a distorting transmission channel;

Fig. 6 is a section from a trellis diagram explaining the Viterbi equalization; and

10 Fig. 7 is a simplified block diagram of a Viterbi decoder, designed for the PSP method.

Description of the Preferred Embodiments:

Referring now to the figures of the drawing in detail and

15 first, particularly, to Fig. 1 thereof, there is illustrated the known principle of a design of a digital transmission system such as is used, for example, in mobile radio for signal transmission.

20

A transmitting device SE accepts an analog source signal Q (generated by a microphone, for example) and leads it to a coder COD. The coder COD comprises, in a way not illustrated, an analog-to-digital converter for digitizing the source

25 signal Q, and can also contain a source coder, a channel coder, an interleaver and a block former which, in a suitable

way, compress the digitized source signal Q, provided with error protection coding, interleave it and subdivide it into data blocks.

5   The coder COD outputs a digital data signal which comprises a symbol sequence $\{d_n\}$ of the data symbols d_0, d_1, ..., which are based, for example, on the value set $\{-1, 1\}$. The symbol sequence $\{d_n\}$ is fed to a modulator device MOD which modulates a high-frequency carrier as a function of the symbol sequence

10   $\{d_n\}$. The modulated, time-dependent, real transmitted signal s(t) calibrated in this case is fed into a transmission channel, that is to say emitted as radio signal via a transmit antenna SA, for example.

15   The data symbols d_0, d_1, ... are denoted below as input data signals (for the modulator device).

Signal distortions and signal interference, both of which depend on the type of transmission channel, can occur during

20   transmission of the transmitted signal s(t) via the transmission channel.

The distorting influence of the transmission channel is described by a channel pulse response $h(\tau, t)$. The additive

25   interference component superimposed on the distorted signal is described by means of a function n(t). A time-continuous

received signal z(t) received at the receiver EM is thereby

yielded as

$$z(t) = \int_0^\infty h(\tau, t) \cdot s(t - \tau) d\tau + n(t) \qquad (1)$$

5

The receiver EM receives the (distorted and interfered) real

received signal z(t) present at the output of a transmission

channel, for example via a receive antenna EA. The received

signal z(t) is fed to a demodulator DMOD. The latter

10 demodulates the received signal z(t). Ready at the output of

the demodulator DMOD is a symbol sequence $\{\hat{d}_n\}$ whose elements

$\hat{d}_n$ are estimated values of the associated input data symbols

d_n. The decoding of the demodulated symbol sequence $\{\hat{d}_n\}$ is

performed in a decoder DCOD arranged on the output side of the

15 demodulator DMOD.

In order to explain the CPFSK modulation, Fig. 2 shows the

structure of a known CPFSK modulator device MOD in the form of

a block diagram. Here, Fig. 2 indicates real variables by a

20 single arrow and complex variables by a double arrow.

In the signal path, the CPFSK modulator MOD has a filter, a

phase accumulator P, a device KE for generating a complex

envelope e(t), a mixing stage MI and a device Re for forming

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the real part of an incoming signal. The (real) modulated

signal s(t) is available at the output of the last named

device Re.

5 Let the filter F have the real pulse response g(t). The pulse

response g(t) is designated below as baseband pulse. The pulse

shape and the (temporal) length of the base and pulse g(t)

define the CPFSK-modulation.

10 It is known that the various pulse shapes, for example square-

wave pulses, $\cos^2$ pulses, or else Gaussian pulses can be used

as baseband pulse.

The pulse length T·K of the baseband pulse g(t) (that is to say

15 the time duration over which the baseband pulse g(t) has a

non-vanishing value) can extend over K = 1, 2, 3, .. symbol

time durations T. Full response types of modulation are spoken

of in the case of K = 1. If the baseband pulse g(t) extends

over a plurality of symbol time durations (that is to say K =

20 2, 3, ..), what are known as partial response types of

modulation are generated.

The real pulse-amplitude-modulated frequency signal f(t)

occurring at the output of the filter F is a linear

25 superimposition of time-shifted versions of g(t) weighted with

the symbols d_n of the input data symbol sequence $\{d_n\}$:

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$$f(t) = \sum_n d_n g(t - nT) \tag{2}$$

The pulse-amplitude-modulated frequency signal f(t) is

5   integrated in the phase accumulator P to form the phase signal

φ(t). In the device KE for generating the complex envelope

e(t), the latter is calculated form the phase signal φ(t) in

accordance with the equation

10   $$e(t) = \exp\{j(\varphi(t) + \varphi 0)\} \tag{3}$$

Here, j denotes the imaginary unit and φ_0 an integration

constant. Consequently, the complex envelope e(t) is yielded

in accordance with:

15

$$e(t) = \exp\left\{ j\left(2\pi\Delta F \int_{-\infty}^{t} f(\tau)d\tau \right) \right\} \tag{4}$$

Here, ΔF denotes the frequency deviation occurring during

modulation. The modulation index η is defined in the usual way

20   by η = 2ΔFT.

The complex envelope e(t) is then multiplied in the mixer

stage MI by the high-frequency carrier of the frequency f_0. The

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real part (Re) of the upwardly mixed signal is the modulated

transmitted signal s(t).

In accordance with the already mentioned article by P. A.

5   Laurent, the preliminary nonlinear CPFSK modulation can be

represented approximately as a linear modulation by using

replacement symbols a_n instead of the input data symbols d_n:

$$e(t) \approx \sum_n a_n \cdot C_0(t - nT)$$ (5)

10

There exist between what is termed the elementary pulse $C_0(t)$

and the baseband pulse g(t) a known, functional relationship

which permits the associated elementary pulse $C_0(t)$ to be

determined for a given type of CPFSK modulation (that is to

15   say for a given baseband pulse g(t)). Reference may be made to

the already mentioned article by P. A. Laurent in this regard.

The following relationship exists between the replacement

symbols a_n and the input data symbols d_n:

20

$$a_n = \exp\left\{ j\pi\eta \sum_{i=-\infty}^{n} d_i \right\}$$ (6)

Each replacement symbol therefore results from the accumulator

input data symbols.

Equation (6) shows that the finite number of replacement

symbols exists only for the case of a rational modulation

index η. Also following from equation (6) is the relationship:

5

$$a_n = a_{n-1} \exp\{j\pi\eta d_n\} \tag{7}$$

Fig. 4a shows a vector or signal space representation of the

possible replacement symbols a_n in the case of CPFSK with a

modulation index η = 0.5. The real part is plotted on the x-

axis, and the imaginary part is plotted on the y-axis. It is

clear that four states exist in the case of η = 0.5, that is

to say the replacement symbols a_n (given an initial phase of 0)

can assume only four values 1, j, -1, -j for d_n = 1 or -1.

These four values a_n are denoted as modulation alphabet.

Let a_{n-1} = 1 by way of example. The state transition relating

to a_n = j and illustrated by the dashed arrow is generated by

an input data symbol d_n = 1. The value of the replacement

symbol relating to the time step n describes the state of the

20 transmitter. It is determined by the sequence of the input

data symbols.

The basic structure, known per se, of a demodulator DMOD for

demodulating a CPFSK signal is firstly explained below with

25 the aid of Fig. 3 for the case of an ISI-affected transmission

channel with additive noise. The demodulator DMOD is to
determine the input data symbol sequence $\{d_n\}$ from the received
signal $z(t)$. Complex variables are represented, in turn, by a
double arrow.

5

The demodulator DMOD can comprise a mixing stage MI which is
operated at the frequency f'_0 ($f_0 = f'_0$ holding ideally), and
which serves to mix the received signals $z(t)$ downward into
the baseband. However, the received signal $z(t)$ can also be
10 processed in the bandpass region at a suitably selected
intermediate frequency.

A lowpass filter TF is used for bandwidth limitation. After
filtering, the downwardly mixed, filtered received signal $z(t)$
15 is sampled with the aid of an analog-to-digital converter ADC.
The sampling is performed at least with the symbol clock $1/T$.

A sequence $\{y_n\}$ comprising complex samples y_n is ready at the
output of the ADC.

20

The sequence $\{y_n\}$ of complex samples are fed to a Viterbi
equalizer VIT. The Viterbi equalizer VIT comprises a channel
estimator KS and an ACS (Add-Compare-Select) calculating unit
ACS. The complex sampling symbols y_n are input into both.

25

In a simple model, see Fig. 5, wherein the transmission channel is described as a finite, clocked automatic machine, the samples y_n can be expressed in accordance with

$$y_n = \sum_{l=0}^{L} a_{n-l} h_l + n_n \tag{8}$$

as a time-discrete convolution of the current and the last L transmitted replacement symbols with L+1 channel pulse responses h_0, h_1, .., h_L, plus an interference symbol sequence $\{n_n\}$, comprising interference symbols n_n, representing the additive interference $n(t)$. The automatic machine comprises a shift register SR consisting of L memory cells T. Located upstream and downstream of each memory cell T in each case are taps (a total of L+1 items) which lead to multipliers M which multiply the values of the data symbols by a channel pulse response h_0, h_1, .., h_L in each case. The output stage of the automatic machine is implemented by an adder AD which adds up the outputs of the L+1 multipliers.

The replacement symbol a_n is fed on the input side to the shift register SR at the nth time step. At this moment, the memory cells of the shift register SR store the previously input replacement symbols $(a_{n-L}, .., a_{n-2}, a_{n-1})$. With each time step, there is generally change in the memory contents of each

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memory cell, since the stored replacement symbol is shifted

into the next memory cell T.

As regards the receiver-end reconstruction of a replacement

5 symbol sequence $\{a_n\}$ comprising p+1 elements (p being a

positive integer), MLSE (Maximum-Likelihood Sequence

Estimation) state that one $\{\hat{a}_n\}$ of the possible (p+1)-element

replacement symbol sequences $\{a_n\}$ whose symbols, weighted with

the channel pulse responses $\hat{h}_0$, $\hat{h}_1$, .., $\hat{h}_L$, and the smallest

10 Euclidean distance from the sequence of the measured samples y_n

is to be determined in the receiver as sequence conjectured to

have been transmitted. The condition for the (p+1)-element

replacement symbol sequence $\{\hat{a}_n\}$ for which a search is being

made is therefore:

$$\sum_{n=0}^{L+p}\left(y_n - \sum_{l=0}^{L} a_{n-l}\hat{h}_l\right)^2 = Minimum \text{ for } \{a_n\} = \{\hat{a}_n\} \qquad (9)$$

The basic idea of the Viterbi algorithm (VA) consists in

recursively solving the MLSE condition equation (9).

20

The memory contents of the shift register SR, and an

accumulated initial phase $\qquad$ are used in

$$\phi_{n-L-1} = \pi\eta \sum_{l=-\infty}^{n-L-1} d_l \qquad \text{the case of}$$

the conventional VA demodulation of a CPFSK-modulated signal

in order to define a combined transmitter and channel state in

a trellis state diagram. This state is described by the $(L+1)$-tuple $A_n = (\phi_{n-L-1}; a_{n-L}, .., a_{n-2}, a_{n-1})$ for the replacement symbols $a_{n-L}, .., a_{n-2}, a_{n-1}$ whilst occurring. This state is unknown in the receiver.

5

In general, the combined transmitter and channel state can be described in the time step n by the $(L+1)$-tuple $Z_n = (\widetilde{\phi}_n, z_n^{L-1}, .., z_n^1, z_n^0)$, the memory contents of the input-end first memory cell T in the time step n being denoted by the variable z_n^0, the memory contents of the downstream memory cells T in the time step n being denoted by the variables $z_n^1, .., z_n^{L-1}$ and the initial phase ϕ_{n-L-1} being denoted by the variable $\widetilde{\phi}_n$. Each of the variables $z_n^{L-1}, .., z_n^1, z_n^0$ can assume the values of the modulation alphabet (that is to say the values of the replacement symbols in the case of the conventional VA).

The transmitted sequence $\{a_n\}$ determines a path through the trellis diagram. In the trellis diagram all possible states Z_n, or a suitable selection of possible states (reduced trellis diagram) are plotted against the time step n in each case. The VA determines the sequence $\{A_n\}$ of the states by means of the trellis diagram in order to estimate the transmitted sequence $\{a_n\}$. The path, determined by means of the sequence $\{A_n\}$,

through the trellis diagram is denoted as best or "shortest" path through the trellis diagram.

Fig. 6 shows a detail of the trellis diagram for the two-time
5 steps n and n+1 with reference to the example of an M-stage replacement symbol data signal and a rational modulation index (M = 4 was selected for the representation in accordance with Fig. 4a). Each circle represents one of the states taken into account in the trellis diagram considered. The states relating
10 to the time step n are denoted by z_n^1, z_n^2, .., in general by z_n^i. A corresponding notation is used for the possible states relating to the time step n+1.

Each state relating to the time step n+1 can be reached by M
15 transitions starting from M different precursor states relating to the time step n. Consideration is now given to those M possible precursor states relating to the time step n which lead to a specific state Z_{n+1}^q, denoted by the index q, in relation to the time step n+1. For each of these M possible
20 precursor states Z_n^{i1}, Z_n^{i2}, .., Z_n^{iM} relating to the time step n, which are illustrated in Fig. 6 in bold (that is to say it is the case in Fig. 6 that i1=1, i2=3, i3=4 and iM=i4=6), the shortest path $P(Z_n^{i1})$, $P(Z_n^{i2})$, $P(Z_n^{i3})$ and $P(Z_n^{iM})$ leading to this state where already determined during the preceding recursion.
25 As already mentioned, these shortest paths are represented in

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each case by the sequence of the precursor states lying on the path, that is to say $P(Z_n^i) = \{Z_0^a, Z_2^b, Z_3^c, .., Z_n^i\}$, a, b, c being the indices of the corresponding states.

5   The question is which of these M paths, which lead to the possible precursor states $Z_n^{i1}, Z_n^{i2}, .., Z_n^{iM}$ is the one, which builds up the shortest path $P(Z_{n+1}^q)$ to this state Z_{n+1}^q when it is continued to the considered state Z_{n+1}^q in the time step n+1.

10   In order to answer this question, a transition metric value (or metric increment) $I(Z_n^{i1}, y_n)$, $I(Z_n^{i2}, y_n)$, .., $I(Z_n^{iM}, y_n)$ is calculated in the case of the VA for each of the transitions considered (between one of the possible precursor states $Z_n^{i1}, Z_n^{i2}, .., Z_n^{iM}$ and the target state Z_{n+1}^q considered),

15   specifically in accordance with:

$$I\left(Z_n^i, y_n\right) = \left(y_n - \sum_{l=1}^{L} z_n^{l-1,(i)} \hat{h}_l - a_n^{(i \to q)} \hat{h}_0 \right)^2 \tag{10}$$

i = i1, i2, .., 1M.

20

Using the notation already introduced, in this case the M possible, known precursor states are described in each case by an (L+1)-tuple $Z_n^i = (\tilde{\varphi}_n;\ z_n^{L-1,(1)}, .., z_n^{1,(i)}, z_n^{0,(i)})$, i = i1, i2, .., iM,

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and $a_n^{(i \to q)}$ denotes that replacement symbol which transfers the precursor state with index i into the target state with index q, that is to say it holds that $z_{n+1}^{(0(q))} = a_n^{(i \to q)}$.

5　Because of the recursive mode of calculation, a minimal metric $Me(z_n^{i1})$, $Me(z_n^{i2})$, .., $Me(z_n^{iM})$ has already been calculated relative to the time step n for each of the M possible precursor states z_n^{i1}, z_n^{i2}, .., z_n^{iM}. The path decision process for the time step n is carried out on the basis of these known M minimal metrics

10　$Me(z_n^{i1})$, $Me(z_n^{i2})$, .., $Me(z_n^{iM})$ of the possible precursor states and the calculated M transition metric values $I(z_n^i, y_n)$ for the respective transitions. It comprises three steps:

The M candidates, denoted by $me^i(z_{n+1}^q)$, for the minimum metric of

15　the target state z_{n+1}^q considered are calculated by means of an addition step ("ADD") as the sum respectively of the minimum metric of one of the precursor states, and of the associated metric increment, doing so in accordance with

20　$$me^i\left(Z_{n+1}^q\right) = Me\left(Z_n^i\right) + I\left(Z_n^i, y_n\right), \quad i = i1, i2, .., iM \tag{11}$$

That one of the M calculated metrics $me^i(z_{n+1}^q)$ which has the smallest value is determined by means of a comparison step

("COMPARE"). This becomes the minimal metric $Me(z_{n+1}^q)$ of the

target state z_{n+1}^q considered.

5   That precursor state of the M possible precursor states

z_n^{i1}, z_n^{i2}, .., z_n^{iM} which is the initial point of the transition to

the state z_{n+1}^q with the minimum metric $Me(z_{n+1}^q)$ is selected by

means of a selection step ("SELECT"), that is to say the index

i, denoted as i(q) is determined for which it holds that

$me^i(z_{n+1}^q) = Me(z_{n+1}^q)$. The correct precursor state is thereby

10   determined. The corresponding path is lengthened into the

state z_{n+1}^q, while the paths, which lead to the remaining

precursor states are not further tracked.

These three steps are known as an ACS (Add-Compare-Select)

15   operation. In descriptive terms, the path from which the

previously received data with the highest probability come is

established in the course of an ACS operation for each state

relative to the time step n+1. Only this path ("survivor") is

progressed, that is to say still taken into account at the

20   next ACS operation. The discarding of possible paths, which is

practiced in each step is the basis of the capability of the

VA to solve MLSE in terms of outlay.

This known application of the VA in demodulating a CPFSK

25   signal is possible only for rational modulation indices. In

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the case of an irrational modulation index, an infinitely large number of states per time step would result, for which reason the trellis could not be realized and implemented.

5   The TST demodulation according to the invention of a CPFSK signal is not based on any prescribed replacement symbol modulation alphabet, but a replacement symbol (transmitter state) one time step back (the (n-1)th one) occurring during the modulation is estimated at the receiver end in the

10   subsequent (nth) time step. The transmitter state is "corrected" in the receiver.

This mode of procedure permits the MLSE of a CPFSK-modulated signal to be carried out on the basis of combined transmitter

15   and channel states, which no longer contain the accumulated phase. The states are allocated such that only the last L input data symbols are taken into account in accordance with the ISI present. Consequently, the ith state of the trellis diagram on which the MLSE according to the invention is based

20   is described in general by the L-tuple $Z_n^i = (z_n^{L-1,(i)}, \ldots, z_n^{1,(i)}, z_n^{0,(i)})$, the variables $z_n^{L-1,(i)}, \ldots, z_n^{1,(i)}, z_n^{0,(i)}$ now being capable of assuming the possible values of the input data symbols d_n, i.e. $\{1, -1\}$.

The actual channel state, unknown in the receiver, relating to the time step n is given by the L-tuple $D_n = (d_{n-L}, \ldots, d_{n-2}, d_{n-1})$.

5   Carrying out an ACS operation in the way according to the invention is explained below for the example of $L = 2$. $\hat{r}_n^{(1)}$ denotes the undisturbed, reconstructed signal relating to the time step n in the state with the index i, that is to say z_n^i. In accordance with equation (9), it holds the:

10

$$\sum_{l=0}^{L} a_{n-l}\hat{h}_l \overset{L=2}{=} a_n\hat{h}_0 + a_{n-1}\hat{h}_1 + a_{n-2}\hat{h}_2$$
$$= a_{n-1}e^{j\pi\eta d_n}\hat{h}_0 + a_{n-1}\hat{h}_1 + a_{n-1}e^{j\pi\eta(-d_{n-1})}\hat{h}_2$$

that is to say

$$
\begin{aligned}
\hat{r}_n^{(i)} &= \sum_{l=0}^{L} a_{n-l}^{(i)}\hat{h}_l \\
&= a_{n-1}^{(i)}e^{j\pi\eta\,d_n^{(i\to q)}}\hat{h}_0 + a_{n-1}^{(i)}\hat{h}_1 + a_{n-1}^{(i)} e^{j\pi\eta\left(-z_n^{0,(i)}\right)}\hat{h}_2 \\
&= a_{n-1}^{(i)}\left(e^{j\pi\eta\,d_n^{(i\to q)}}\hat{h}_0 + \hat{h}_1 + e^{j\pi\eta\left(-z_n^{0,(i)}\right)}\hat{h}_2\right)
\end{aligned}
\tag{12}
$$

15

$a_{n-1}^{(i)}$ denoting the replacement symbol assigned to the ith transmitter state, and $d_n^{(i\to q)}$ denoting that input data symbol which in relation to the time step n leads from the precursor

20   state with index i into the target state with index q. Since

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the state description in the case of the MLSE according to the

invention no longer is performed with the aid of the

replacement symbols, $a_{n-1}^{(i)}$ is unknown a priori in the receiver

(in the case of the previously described conventional VA, $a_{n-1}^{(i)}$

5    $= z_n^{0,(i)}$ would be prescribed by the output state considered).

An estimated value for $a_{n-1}^{(i)}$ is therefore determined in the

receiver. The estimated value is denoted as $\hat{a}_{n-1}^{(i)}$. The principle

of facilitating an MLSE on the basis of a state description by

10   means of input data symbols by determining an estimated value

$\hat{a}_{n-1}^{(i)}$ for $a_{n-1}^{(i)}$ is a fundamental aspect of the TST.

After the determination of an estimated value $\hat{a}_{n-1}^{(i)}$ for $a_{n-1}^{(i)}$, the

undisturbed, reconstructed signal is used in accordance with

15   equation (12) together with the (disturbed) received signal y_n

to calculate a transition metric value

$$I\left(Z_n^i, y_n\right) = \left(y_n - \hat{r}_n^{(i)}\right)^2 \tag{13}$$

20   The determination of the estimated value $\hat{a}_{n-1}^{(i)}$ can be performed

in a simple way on the basis of the estimated value $\hat{a}_{n-2}^{(i)}$

determined for the previous time step: the transmitter state

is corrected, as it were.

A first possibility for this consists in assuming by way of simplification that $a_{n-1}^{(i)}$ is identical for all states i (in the same time step), that is to say only a single "generalized" transmitter state need be estimated for each time step. This can be performed in accordance with

$$\hat{a}_{n-1}^{(i)} = \hat{a}_{n-1} = \hat{a}_{n-2} \exp\left\{j\pi\eta\hat{d}_{n-1}\right\} \tag{14}$$

$\hat{d}_{n-1}$ being a (provisional) value, generated by means of a DFE technique, for example, for the reconstructed input data symbol d_n. An equalizer with quantized feedback, what is termed DF (Decision Feedback) equalizer, is suitable for carrying out such a method.

A second possibility for calculating the estimated values $\hat{a}_{n-1}^{(i)}$ for $a_{n-1}^{(i)}$ from $\hat{a}_{n-2}^{(i)}$ is yielded by the relationship:

$$\hat{a}_{n-1}^{(i)} = \hat{a}_{n-2}^{(i)} \exp\left\{j\pi\eta\hat{d}_{n-1}\right\} \tag{15}$$

Again, the DFE technique can be used to determine a value for $\hat{d}_{n-1}$.

A third possibility consists in determining the estimated values $\hat{a}_{n-1}^{(i)}$ on the basis of the already mentioned PSP method.

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The PSP method is described in US 5,432 821 and is incorporated in the present document by reference. The PSP method is based in general on generating an estimated value for a variable unknown at the receiver end and assigned to a

5   specific state on the basis of information (already present in the receiver) on the path, which leads to this state.

The determination of $\hat{a}_{n-1}^{(i)}$ using the PSP method can therefore be described in general form by the equation

$$\hat{a}_{n-1}^{(i)} = G\left(S\left(Z_n^i\right)\right) \tag{16}$$

Here, $S(Z_n^i)$ is the sequence, assigned to the path $P(Z_n^i)$ (that is to say that path P which ends in the state in relation to the

15   time step n) of already decided input data symbols which is denoted below by

$$S\left(Z_n^i\right) = \left(d_0^{P\left(Z_n^i\right)}, d_1^{P\left(Z_n^i\right)}, \ldots, d_{n-1}^{P\left(Z_n^i\right)}\right) \tag{17}$$

20   G($\cdot$) denotes an estimator which assigns an estimated value to such a sequence of decided input data symbols.

A more special form of the PSP method for determining $\hat{a}_{n-1}^{(i)}$ is specified by the expression

25

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$$\hat{a}_{n-1}^{(i)} = \hat{a}_{n-2}^{(i)} \exp\left\{ j\pi\eta G\left(S\left(Z_n^i \right) \right) \right\} \tag{18}$$

A simple example for $G(\cdot)$ in this case is:

5 $\quad G\left(S\left(Z_n^i \right) \right) = d_{n-1}^{P\left(Z_n^i \right)} = z_n^{0,(i)} \tag{19}$

That is to say

$$\hat{a}_{n-1}^{(i)} = \hat{a}_{n-2}^{(i)} \exp\left\{ j\pi\eta d_{n-1}^{P\left(Z_n^i \right)} \right\} \tag{20}$$

10   The PSP method can be used to determine further parameters

unknown in the receiver. Reference is made in this connection

to the fact that the application of the TST demodulation

method described above depends sensitively on how accurately

the modulation index η of the transmitter is known in the

receiver. The modulation index used in the receiver for

demodulation is denoted below by $\hat{\eta}$. Slight deviations between

the modulation index η and the modulation index $\hat{\eta}$ set at the

receiver end are enough in themselves to lead to substantial

20   losses in the efficiency of the TST algorithm. The reason for

this is that a deviation $\Delta\eta$ in the modulation index assumed at

the transmitter end with respect to the true modulation index

η (that is to say $\eta = \hat{\eta} + \Delta\eta$) in relation to the time step n

set a phase error $\Delta\phi_n^{(\eta)}$ in accordance with the relationship

$$\Delta \phi_n^{(\eta)} = \pi \Delta \eta \sum_{k=0}^{n} d_k \qquad\qquad (21)$$

The phase error $\Delta \phi_n^{(\eta)}$ is therefore a function of the

5   transmitted data sequence $(d_0, d_1, .., d_{n-1})$. In accordance with

equation (12), this phase error acts on the undisturbed,

reconstructed signal.

For the case wherein the frequency modulation is undertaken,

10 as illustrated in Fig. 2, relative to a carrier frequency f_0,

and at the receiver end the shift by the carrier frequency is

nullified by mixing downward, a frequency offset can occur

between the carrier frequency f_0 of the transmitter and a

frequency f'_0 which is used for mixing downward. This frequency

15 offset, or else a drift one of the two frequencies, likewise

act as phase error in the undisturbed, reconstructed signal.

For example, a constant frequency offset $\Delta f = f_0 - f'_0$ leads to a

phase error of a linear profile:

20    $\Delta \phi_n^{(\Delta f)} = 2\pi \cdot \Delta f \cdot n \cdot T \qquad\qquad (22)$

The phase between the reconstructed signal (see equation 12)

and the receive signal y_n is determined in accordance with

$$\hat{\phi}_n^{(i)} = -\arg\left(\hat{r}_n^{(i)} \cdot y_n^*\right) \tag{23}$$

in order to determine the phase error. In this case, "arg"
denotes the angle of a complex number in radian measure and
5 "*" denotes the complex conjugate with the complex number
provided with the asterisk.

The phase error $\hat{\phi}_n^{(i)}$ is used in the following way for phase
correction of the undisturbed, reconstructed signal $\hat{r}_n^{(i)}$:

$$\hat{a}_{n-1,corrected}^{(i)} = G\left(S\left(Z_n^i\right)\right) \cdot \exp\left(j\hat{\phi}_n^{(i)}\right) \tag{24}$$

that is to say Gl. 12

$$\hat{r}_{n,corrected}^{(i)} = \hat{r}_n^{(i)} \cdot \exp\left(j\hat{\phi}_n^{(i)}\right) \tag{25}$$

being yielded in accordance with equation 12.

A phase error $\hat{\phi}_n^{(i)}$ likewise forms an unknown parameter in the
20 VA. A data based estimate of this unknown parameter can
likewise be realized with the PSP method. That is to say for
each path ("survivor") within the trellis diagram the input
data symbol decisions made for this path are used to estimate
this unknown parameter by analogy with equations 16, 17. The

unknown parameter is estimated after each time step, and

updating of it is performed. The updated estimated value of

the unknown parameter (phase error $\hat{\phi}_n^{(i)}$) is then used, as

already described, for the next calculation of a transition

5   metric value.

In order to obtain estimated values for the phase errors, the

phase errors, which were determined for the individual states,

are filtered. It is possible, for example, to use as filter an

10 IIR filter of first order, the filtered estimated value for

the state with the index i in relation to the nth time stamp

being yielded in accordance with the following relationship:

$$\theta_n^{(i)} = \beta_1 \hat{\phi}_n^{(i)} + \beta_2 \hat{\phi}_{n-1}^{(i)} \tag{26}$$

β_1 and β_2 denoting suitably selected filter constants in this

case.

Fig. 7 shows a block diagram of a device according to the

20 invention for carrying out the TST-MLSE. The PSP method is

used in this example both to determine the first estimated

values (for the replacement symbols) and to determine the

second estimated values (for the phase errors).

As illustrated in Fig. 3, the Viterbi equalizer VIT accepts the complex samples y_n and outputs the estimated input data symbols $\hat{d}_n$ with a demodulation delay D. On the one hand, the equalizer VIT comprises a section A2 for carrying out the ACS operations. The section A2 comprises a first unit ADD for calculating the metric values, a second unit COM, wherein a comparison of the updated metrics is undertaken, and a third unit SEL, which carries out the selection operation, that is to say progresses that path into the target state considered for which the smallest metric has been determined in the second unit COM.

On the other hand, the equalizer VIT comprises a first calculating section A1, which is connected upstream of the second calculating section A2 in the signal path. The first calculating section A1 comprises a channel estimator KS, M unknown-parameter estimators UPE_1, .., UPE_M, M calculating sections TM_1, .., TM_M for calculating transition metric values, and a path memory MEM.

The channel estimator KS calculates the estimated channel parameters $\hat{h}_0$, .., $\hat{h}_L$ and communicates them to the calculating sections TM_1, .., TM_M for calculating the transition metric values.

Sequences of decided input data symbols $S(z_n^i)$ with reference to

all paths tracked in the trellis diagram are stored in the

path memory MEM. Their number corresponds to the state

considered in the trellis diagram, that is to say is equal to

5   M or greater. The sequences of decided input data symbols are

updated in relation to each time step.

The mode of operation of the Viterbi equalizer VIT is

considered below with reference to the prescribed target state

10   z_{n+1}^q. The unknown-parameter estimators UPE_1, .., UPE_M are fed

from MEM over sequences of decided input data symbols assigned

to the possible precursor states of the state considered. The

unknown-parameter estimators UPE_1, .., UPE_M estimate for each

path of the state $\hat{a}_{n-1}^{(i1)}$, .., $\hat{a}_{n-1}^{(iM)}$ of the transmitter relative to

15   a preceding time step, and communicate the estimated values to

the calculating sections TM_1, .., TM_M for the calculation of

the transition metric values. The estimated values for the

replacement symbols (that is to say the path-referred

transmitter states) are determined in the tracking mode in

20   accordance with equation 18.

The transition metric values $I(z_n^{i1}, y_n)$, .., $I(z_n^{iM}, y_n)$ calculated

in the calculating sections TM_1, .., TM_M are fed to the

addition unit ADD and further processed in the already

25   described way in the units COM and SEL. After the ACS

operations have been carried out relative to all target states
relating to the time step n+1, the paths (survivors) are
determined for this time step.

5   The Viterbi equalizer VIT can also be designed in terms of
circuitry (in a way not illustrated) such that the ACS
operations are carried out in parallel for all target states.

In order to estimate the phase errors of the undisturbed,
10   reconstructed signals, further unknown-parameter estimators
UPE'_1, .., UPE'_M are present which undertake an estimation of
the respective phase errors in accordance with equations 21
and 23 and/or 26. The estimated phase errors or estimated
filtered phase errors are communicated to the calculating
15   sections TM_1, .., TM_M. The efficiency of the TST method is
substantially raised by taking account of these phase errors
(in accordance with equations 24 and 25, similarly in the case
of the fundamental use of filtered estimated phase errors) in
the calculation of the transition metric values in the
20   calculating sections TM_1, .., TM_M.

It may be pointed out in conclusion that it is also possible
to provide a DF equalizer as an alternative to the Viterbi
equalizer VIT, shown in Fig. 7, which determines unknown
25   parameters on the basis of a PSP method.